CHAPTER 11.4
OPERATIONAL AMPLIFIERS

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DIRECT-COUPLED AMPLIFIERS

Samuel M. Korzekwa

A direct-coupled amplifier has frequency response that starts at zero frequency (dc) and extends to some specified upper limit. To obtain the zero-frequency capability, such amplifiers are normally direct-coupled throughout; i.e., they do not use capacitive or transformer coupling (except for auxiliary higher-frequency compensation or signal transmission).

The primary sources of error of a direct-coupled amplifier are initial offset, drift, and gain variations, errors usually dependent on temperature, aging, and so forth. The gain-variation problem can be minimized by feedback gain-stabilization techniques, but offset and drift errors are usually not handled so effectively by feedback techniques. A bias shift or drift error cannot be distinguished from a signal response because their output responses are identical. Various techniques are available to minimize this drift problem, and several different methods may be used in an amplifier.

One method of minimizing drift is to use a balanced topology, as in differential amplifiers, whereby the drift errors tend to cancel out. A more complicated but effective method is the modulated-carrier-amplifier approach: the signal is first converted into a carrier signal, amplified using an ac-coupled amplifier, and then demodulated to a baseband signal. The function performed by the modulated-carrier amplifier is identical with that of a direct-coupled amplifier, but the signal processing technique differs.

An example of a direct-coupled transistor amplifier is illustrated in Fig. 11.4.1. The primary signal path, shown in heavy lines, directly couples the input signal to the amplifier output. This configuration also provides gain and bias stabilization via the direct-coupled R_3 feedback path. Usually the low-frequency amplifier voltage gain is primarily determined via the R_1 and R_3 components, while the function of the C_1 capacitor and R_6 resistor is to provide high-frequency compensation or stabilization. Sources of error for this topology include the initial V_{BE} offset of the Q_1 input transistor and the subsequent drift caused by the temperature dependence of this offset.

Differential Amplifiers

A differential amplifier is a dual-input amplifier that amplifies the difference between its two signal inputs. This amplifier may have an output that is single-ended (one output) or it may have a differential output.

The differential amplifier eliminates or greatly minimizes many common sources of error. The drift problem encountered in direct-coupled amplifiers can be handled more effectively by the differential approach. A second major advantage of a differential amplifier is its ability to reject common-mode signals, i.e., unwanted signals present at both of the amplifier inputs or other common points. Common-mode performance is usually a critical requirement of instrumentation amplifiers.

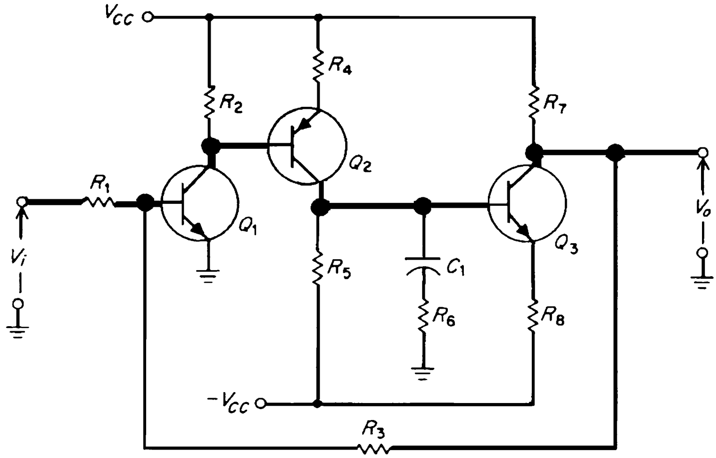


FIGURE 11.4.1 Direct-coupled transistor amplifier.

The basic circuit commonly used in differential amplifiers is shown in Fig. 11.4.2. Such differential pairs can be constructed using separate devices or in integrated-circuit form. The integrated package yields additional advantages since the parameter differences between the units of the integral differential pair are usually much less than if separate devices are used. Thus the units of such integral pairs tend to track differentially more closely, even though their individual parameters may vary in absolute value. Also, many of the passive components in the integrated amplifiers track better. Figure 11.4.3 shows a typical differential transistor amplifier. For further detailed information, including analysis procedures, design techniques, and application data, refer to the literature.

Chopper (Modulated-Carrier) Amplifiers

The chopper amplifier performs the function of a direct-coupled amplifier by using a carrier frequency and an ac-coupled amplifier. The modulated-carrier approach is used specifically to minimize drift and offset types of errors. Although this approach is often more complex to implement, the performance improvement makes this technique desirable for applications demanding low-drift performance.

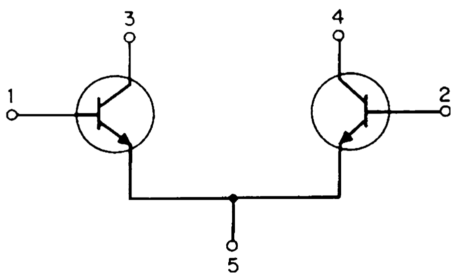


FIGURE 11.4.2 Differential amplifier pair.

Figure 11.4.4 shows the block diagram of a modulated-carrier amplifier. While the input and output signals are dc-coupled, the interstage coupling between the modulator, amplifier, and demodulator may be either a capacitor or a transformer.

Since modulator and demodulator are usually operated synchronously, the carrier delay must be maintained at acceptable levels. Low-pass filters are generally included in the modulator and demodulator. The choice of carrier frequency depends on the application. Typical examples of low-frequency carriers are 60 and 400 Hz. Carrier frequencies above 10 kHz are also feasible. Since the chopper and demodulator generate noise at their carrier frequency and its harmonics, the carrier frequency should normally be chosen above the baseband frequency range of interest. Early chopper modulators and demodulators were mechanical devices, but modern types employ solid-state electronics.

A variation of the modulated-carrier technique uses the chopper-stabilized approach illustrated in Fig. 11.4.5, which includes an additional parallel ac-coupled high-frequency signal path. The two signal paths

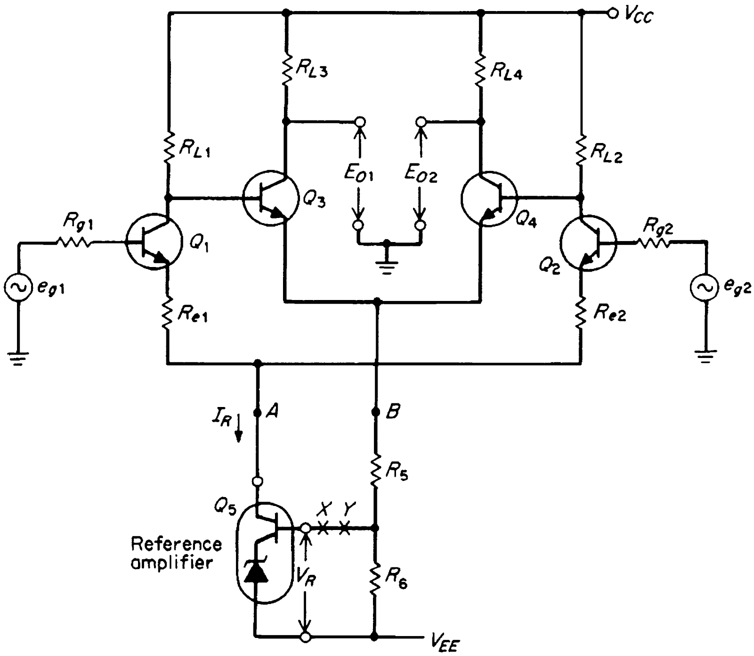


FIGURE 11.4.3 Two-stage differential amplifier with common-mode feedback.

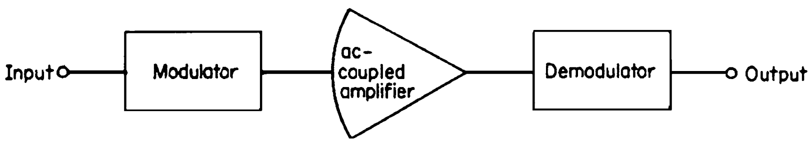


FIGURE 11.4.4 Modulated-carrier (chopper) amplifier.

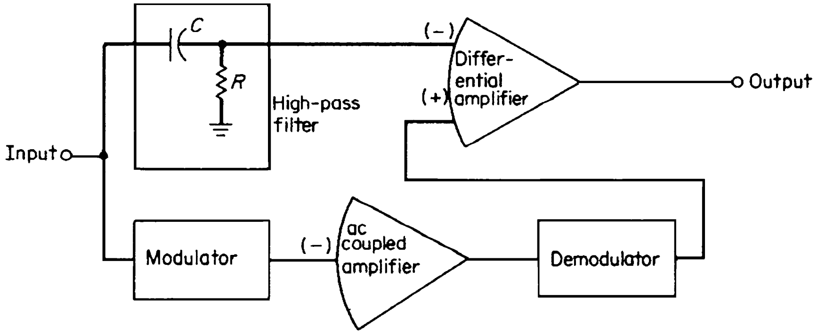


FIGURE 11.4.5 Chopper-stabilized amplifier.

have gains and bandwidths tailored to a crossover frequency; e.g., when the high-frequency signal path becomes dominant, the modulated-carrier signal path ceases to contribute to the sum total. This variation offers the low-drift advantages of the carrier-modulated system with much higher bandwidth.

Operational Amplifiers

An operational amplifier is intended to realize specific signal processing functions. For example, the same operational amplifier (depending on the externally added components) can be used as an integrating amplifier, a differentiating amplifier, an active filter, or an oscillator, among others. Applications of operational amplifiers also include such functions as impendance transformers, regulators, and signal conditioning. They are versatile building blocks that can also be used in nonlinear applications to realize functions such as logarithmic amplifiers, comparators, ideal rectifiers, and so forth.

The early application of operational amplifiers was largely in the area of analog computations. The requirements placed on the amplifiers were severe, and the cost was high. Currently, however, these high-performance functions are available at low cost and are widely used.

An operational amplifier can be either direct-coupled or ac-coupled. Most operational amplifiers have differential inputs and consequently realize common-mode rejection; however, many operational amplifiers are single-ended. Their power capability covers a wide range, and they can function as a power driver. The open-loop bandwidth can range from below 1 kHz up to the megahertz range. Voltage gain can be designed from unity to above 100,000.

To design or use the many excellent models now available commercially, it is necessary to understand the attributes and limitations of these versatile building blocks. The reader is referred to the literature and, of course, to the data sheets and applications literature of individual suppliers of op amps.

Application of Operational Amplifiers

Figure 11.4.6 shows the block diagram for an operational amplifier with associated external elements. A_1 represents the transfer function of the amplifier, either current gain or voltage gain, which is generally frequency-dependent. Z_i and Z_f are the primary elements that normally determine the closed-loop transfer function for this operational amplifier. The indicated Z_{in} and Z_l are equivalent summing node and load impedances, respectively, which are usually factored into the error terms of the composite transfer function. All the elements are general impedances that may be real or complex. Thus the simplified transfer function of this operational amplifier can be written

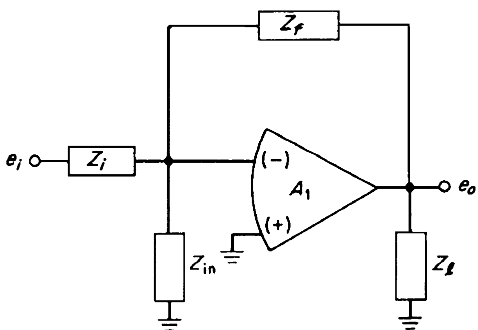


FIGURE 11.4.6 Operational amplifier A_1 with external impedances which determine its functional application.

$$e_o/e_i \approx -Z_f/Z_i \tag{1}$$

The error terms of the complete transfer function equation are not shown in Eq. (1); however, the complete equation with error terms can be readily derived or obtained from the literature. Thus, if the required constraints are adhered to, Eq. (1) can be used to generate various transfer functions, as illustrated in the examples given below.

Integrating Amplifier. An example of an integrating amplifier is shown in Fig. 11.4.7. Assume that the external-component values are as indicated and that the A_1 amplifier is a widely used commercial integrated circuit op amp. Using Laplace transform notation, the integrator transfer function becomes

$$(e_o/e_i)S = (-)1/RCS \tag{2}$$

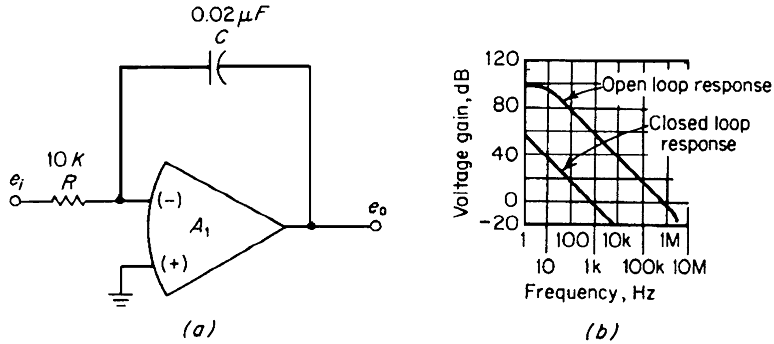


FIGURE 11.4.7 Operational amplifier used for integration: (*a*) basic schematic; (*b*) frequency response.

Using the component values specified in the figure, the integrator transfer function can then be written

$$(e_o/e_i)S = (-)5000/S \tag{3}$$

An alternative frequency-domain representation of the above integrator transfer function can also be used.

$$(e_o/e_i)f = (-)800/jf \tag{4}$$

The closed- and open-loop frequency responses of this operational integrator amplifier are illustrated in Fig. 11.4.7*b*.

The frequency range of application depends primarily on the accuracy desired. In this example, 1 Hz to 10 kHz frequency is considered a realistic range of operation. For very low frequencies the error tends to increase because of inadequate excess gain within the operational amplifier, whereas for high frequencies the closed-loop gain becomes low and drift and offset errors may then become important.

Differentiating Amplifier. An example of a differentiating amplifier is shown in Fig. 11.4.8. The following simplified closed-loop equations are applicable:

$$(e_o/e_i)S \approx (-)RCS \tag{5}$$

$$(e_o/e_i)S \approx (-)0.165S \tag{6}$$

$$(e_o/e_i)f \approx (-)jf \tag{7}$$

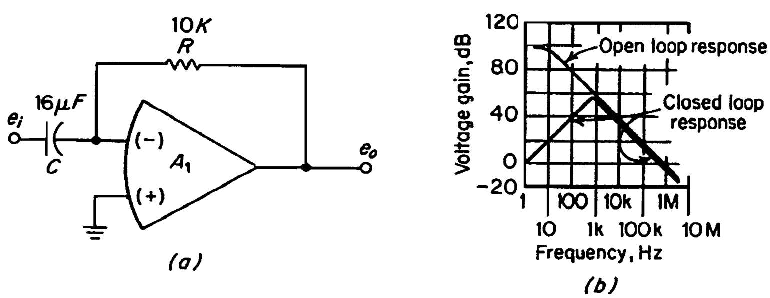


FIGURE 11.4.8 Operational amplifier used for differentiation: (*a*) basic schematic; (*b*) frequency response.

The frequency responses of this operational amplifier differentiator are shown in Fig. 11.4.8*b*. The open-loop response of the operational amplifier of Fig. 11.4.7 is used, and the differentiator closed-loop response is superimposed onto it.

The frequency range of this operational amplifier differentiator depends primarily on the closed-loop differentiation accuracy required. The difference between open- and closed-loop transfer functions (usually referred to as *excess gain*) can be used to predict the accuracy of the function being generated. A realistic frequency range is from 0.01 to 100 Hz. At very low frequencies the closed-loop gain becomes very small, and consequently errors such as drift and offset may become critical. At high frequencies the accuracy degrades, and ultimately an integration function is generated rather than the differentiation function intended.

Servo Amplifiers

The function of a servo amplifier, one of the principal components in a control feedback system, is to amplify the input (usually low-level) error signals and to provide rated drive power to the load (the servo actuator). Servo amplifiers can be dc- or ac-operated, and can be linear or nonlinear.

Direct-Coupled Servo Amplifiers. A direct-coupled servo amplifier can operate on dc error signals, i.e., zero frequency signals. The bandwidth of a dc servo amplifier is often quite large, and thus it has the capability of dynamic operation. In practice it is often operated in the dynamic mode, and the transient-response characterization is often used.

A servo amplifier is intended to drive or control its load to some prescribed reference level. It is the drift and offset errors associated with this control function that are the main concern in this dc mode of operation. Thus, if the actuator is positioned at its exact reference level, the amplifier should provide zero drive power. However, because of initial offsets or subsequent temperature or aging effects, the amplifier may still provide unwanted drive to the load. The usual solution is to minimize these problems by circuit design, e.g., tracking or balanced configurations, or by use of an intermediate-carrier modulated by the dc error input signals, subsequently demodulated for use as the direct-coupled drive to the actuator.

AC Servo Amplifiers. An ac servo amplifier operates at a selected fixed frequency and is consequently a carrier-system amplifier. The most commonly used carrier frequencies are 60 and 400 Hz, but ac servo systems can be designed to operate at almost any frequency compatible with the servo actuator used.

The principal advantage of ac servo amplifiers is that the previously discussed drift and offset problems present in dc servo amplifiers are virtually eliminated. Another advantage is that the prime power from the actuator now can be supplied directly from the line, for example, 60-Hz 115-V source.

For applications wherein the servo actuator or load is a servomotor, the load is often tuned to the carrier frequency via the addition of a capacitor in parallel with the motor. Note that the servomotor impedance is a function of the motor speed; however, the losses (usually resistive) change, whereas the parallel inductance remains virtually constant. Thus tuning the load causes it to become real (resistive), and consequently the efficiency of the amplifier output stage improves significantly.

The output stage of an ac servo amplifier is usually operated class B or AB and in a push-pull configuration to obtain the best possible drive efficiency. Figure 11.4.9 shows an example of a push-pull tuned-load ac servo amplifier.

Nonlinear Servo Amplifiers. This class of servo amplifier uses the load to filter the highly nonlinear drive signals resulting in improved drive efficiencies and higher realizable driver-power capabilities. The drivers essentially act as switches wherein the dissipation losses are low when the switches are either on or off. Since the loads or servo actuators are usually highly reactive, they can be advantageously used in this manner. The drive power can be readily derived from a dc source via a pulse-width-modulated scheme. In addition, an ac source of power can also be used via a phase-modulation technique. Some examples of high-power switching devices used for ac nonlinear servo amplifiers are thyratrons and silicon controlled rectifiers (SCRs).

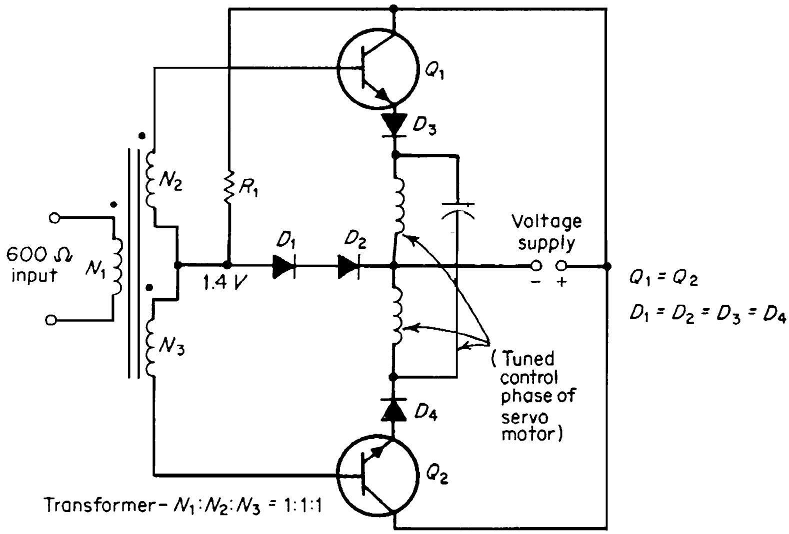


FIGURE 11.4.9 Servo amplifier used as a motor drive.

OPERATIONAL AMPLIFIERS FOR ANALOG ARITHMETIC

Joseph P. Hesler

Analog Multiplier Circuits

Circuits used for analog multiplication fall into three categories: transconductance multipliers, averaging-type multipliers, and exponential multipliers.

Transconductance Multipliers

The most prominent type of transconductance multiplier uses the property of the bipolar transistor; i.e., its collector current and transconductance are linearly related. The balanced transistor differential amplifier used in these circuits offers good accuracy, wide bandwidth, and low cost.

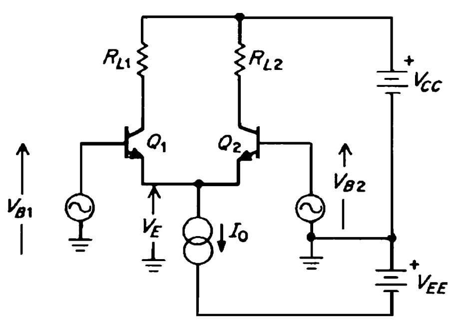


FIGURE 11.4.10 Basic differential amplifier for analog multiplication.

In the differential amplifier as shown in Fig. 11.4.10, the collector currents I_{c1} and I_{c2} are functions of the differential input voltage $V_{b1} - V_{b2}$. It is assumed that the current transfer ratios of the two transistors (Q_1 and Q_2), the junction ambient temperatures, and the base-emitter junction saturation currents are the same, as is probable in integrated-circuit fabrication.

The diode equation governs the relationship between the transistor emitter currents I_{Ei} and the base-to-emitter voltage V_{BE1}

$$I_E = I_S \exp\left(\frac{V_{BE}q}{kT} - 1\right)$$

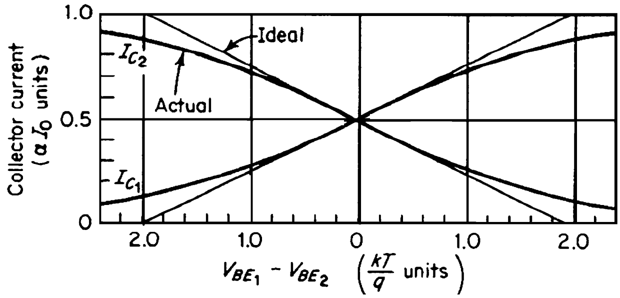


FIGURE 11.4.11 Transfer curves on which analog multiplication is based.

where I_S = junction saturation current
q = electron charge = 1.6×10^{19} C
k = Boltzmann's constant = 1.38×10^{-23} W/s·K
T = junction temperature (K)

The relationships of collector currents, shown graphically in Fig. 11.4.11 are

$$I_{c1} = \frac{\alpha_1 I_0}{1 + \exp[(V_{BE_2} - V_{BE_1})q/kT]} \qquad I_{c2} = \frac{\alpha_2 I_0}{1 + \exp[(V_{BE_1} - V_{BE_2})q/kT]}$$

For zero differential input, $V_{BE_1} = V_{BE_2}$ and $\alpha_1 = \alpha_2$, and so

$$I_{c_1} = I_{c_2} = \alpha I_0/2$$

At zero input, $\Delta V_{BE} = 0$, the maximum single-ended transconductance occurs at

$$gm_{\max} = \alpha I_0/(4kt/q)$$

For a differential output connection the maximum transconductance is twice this value. The linear range of transconductance extends over a differential input signal range of about 50 mV peak to peak at room temperature.

Linear multiplication occurs when one input varies the I_0 term and the other input is used to vary ΔV_{BE}.

$$\Delta I_c = gm\,\Delta V_{BE} = \frac{\alpha I_0}{2kT/q}\Delta V_{BE}$$

The linear range of input voltage can be extended by inserting resistance in series with each emitter. This increase in allowable signal is accompanied by a corresponding reduction in transconductance. For I_0 equal to 2 mA, the addition of 50 X in series with each emitter increases the linear input swing by a factor of 3 and reduces the transconductance to one-third. Optimization of the transconductance multiplier operation requires a trade-off between linearity and error sources because of offset voltages and thermal noise.

A typical four-quadrant multiplier using the differential amplifier as building blocks is shown in Fig. 11.4.12. This circuit is generally used with differential inputs to obtain maximum linearity and common-mode signal rejection.

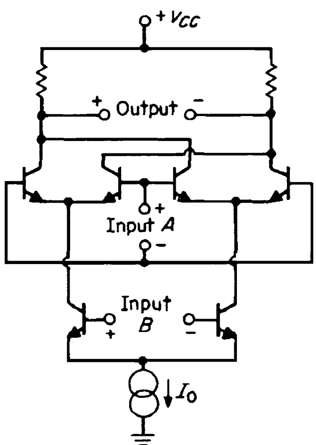


FIGURE 11.4.12 Four-quadrant transconductance multiplier.

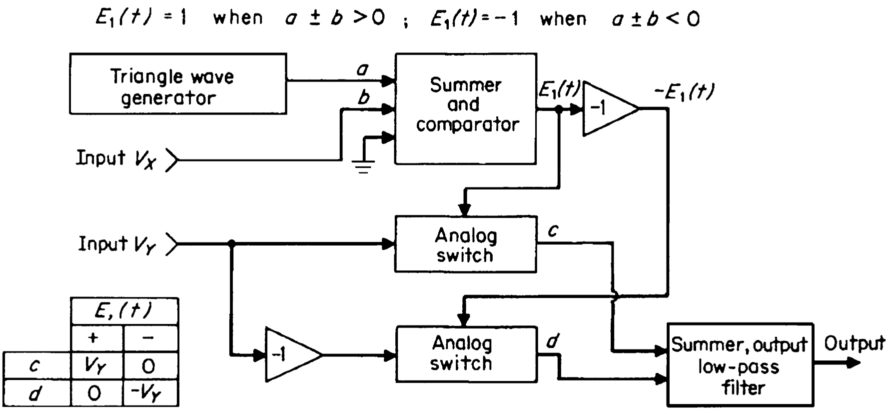


FIGURE 11.4.13 Block diagram of an averaging-type multiplier.

Averaging-Type Multipliers

Two types of averaging multipliers have found extensive use. The first type, the pulse height and width multiplier, uses one of the input variables to modulate the height or amplitude of a pulse train, while the pulse width is modulated with the other input variable. The pulse area, averaged over a suitable period, is proportional to the product of the input variables. High pulse rates permit short averaging intervals and fast response. These multipliers offer very good accuracy and stability but are more expensive than the transconductance multipliers. A block diagram of a typical pulse height and width averaging multiplier is shown in Fig. 11.4.13.

A second version of the averaging multiplier is the triangle averaging multiplier. In this type a high-frequency triangular waveform is generated and combined with the input variables to form an averaged output proportional to the product of the input signals. These multipliers are less accurate than the pulse height and width multipliers and are being displaced by transconductance multipliers.

A third type of averaging multiplier is the time-base multiplier. This approach uses a comparator to sense the time interval required for the integral of a reference input voltage to equal the amplitude of a sample of one input variable. During the same interval, 0 to T, the other input variable is integrated to produce an output V. This type of circuit can be built with a few components to provide moderate accuracy, 1 to 5 percent, at low cost.

Exponential Multipliers

The first electronic multipliers were of the exponential type. In these circuits, resistor-diode networks are designed to provide a current or voltage output approximating the square of the input. These multipliers are also called *quarter-square* multipliers based on the identity

$$XY = {}^1/_4\,[(X+Y)^2 - (X-Y)^2]$$

The piecewise-linear approximations to the squared response result in "lumpy" error characteristics. Also, the amount of circuitry required to compute the quarter-square algorithm is expensive. Although these multipliers are capable of good accuracy and bandwidth, they are becoming obsolete.

A second type of exponential multiplier uses logarithmic amplifiers, a summer, and an antilog amplifier to implement the relationship

$$XY = \text{antilog}_a\,(\log_a X + \log_a Y)$$

Semiconductor diodes are available with excellent logging characteristics over many decades of bias current. These diodes are used with operational amplifiers to realize the necessary functions. The circuits can provide moderate to good accuracy (with thermal compensation) and good bandwidth. Applications are restricted by unipolar input requirements and differential drift because of thermal effects.

Multiplier Error Sources

Offset Error. There are two subclasses of offset error. The first is static offset caused by variances in component parameters such as saturation current I_s, in transistors and diodes. The second error is a result of drift in component parameters. While the initial static offset can be trimmed with external adjustments, the drift components must be compensated by introducing complementary temperature coefficients within the circuit. A common source of drift is local heating of diode junctions and resistive components having nonzero temperature coefficients.

Feedthrough Error. Feedthrough errors result from nonideal transfer characteristics. Two feedthrough error contributions can exist. The first, E_{FY}, is defined as the output owing to the input variable Y when the X input is zero. The second, E_{FX}, is the complementary function owing to an X input. Feedthrough erros may result in dc, fundamental, and harmonic components of the contributing input signal.

Gain and nonlinearity errors. Gain errors produce output deviations from the expected scale factor of the multipliers. Nonlinearity of the transfer functions of the multiplier can produce additional error contributions, as previously discussed. Gain is most apt to vary as a function of the combined input-signal values because the internal components are operated over a range of bias conditions.

In certain multiplier applications, required transient responses may overtax the bandwidth capabilities of the circuits. In these instances additional error terms may appear as a result of limited slew rate of the circuits and differential phase response between the input channels.

Squaring Circuits

Squaring circuits are readily implemented by introducing the variable to both inputs of a multiplier circuit. Alternatively, the resistor-diode squaring circuits used in exponential multipliers may be applied directly.

Dividing and Square-Root Circuits

Division and square-root functions can be implemented with basic multiplier circuits by altering the interconnections. A typical dividing-circuit connection is shown in Fig. 11.4.14. The multiplier output is fed back

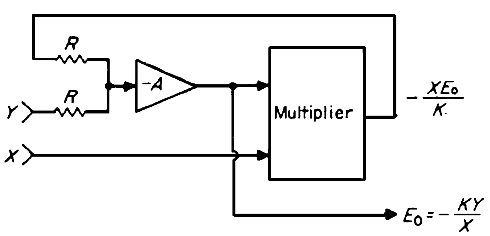
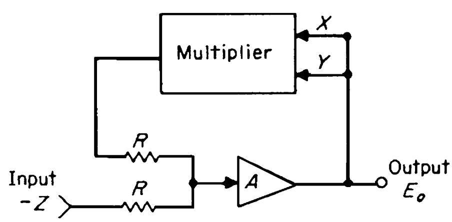


FIGURE 11.4.14 Dividing circuit.



$$-\frac{Z}{R} + \frac{E_o^2}{KR} \cong 0$$

$$E_o^2 = KZ$$

$$E_o = \sqrt{KZ}$$

FIGURE 11.4.15 Square-root circuit.

through a summing amplifier to one of the multiplier inputs. The summing amplifier maintains an equivalence between the numerator Y and the multiplier output XE_0/K.

$$Y/R = -XE_0/KR \qquad \text{or} \qquad E_0 = -KY/X$$

Square-root circuits can be implemented in a similar method with another feedback connection involving a multiplier. In this instance, the output is used as each of the two multiplier-input variables, and the multiplier output is summed with the variable whose square root is desired. The square-root connection is shown in Fig. 11.4.15. The nonnegative input limitations should be stated for division and square-root connections.

LOW-NOISE OPERATIONAL AMPLIFIERS

Clayton R. Roberts

Low-Noise Op-Amp Design

The use of the standard operational-amplifier module or chip has grown into many areas of design because of its low cost, ease of design, and low number of supporting components. Some of the well-known basic amplifier uses are shown in Fig. 11.4.16. Low-noise amplifier developments have moved into the op-amp category for the same reasons. Applications in this area include microphone pre-amplifiers, threshold detectors, instrumentation amplifiers, and sensor amplifiers. Good low-noise op-amps are available, and understanding some of the basic design principles is helpful when using them.

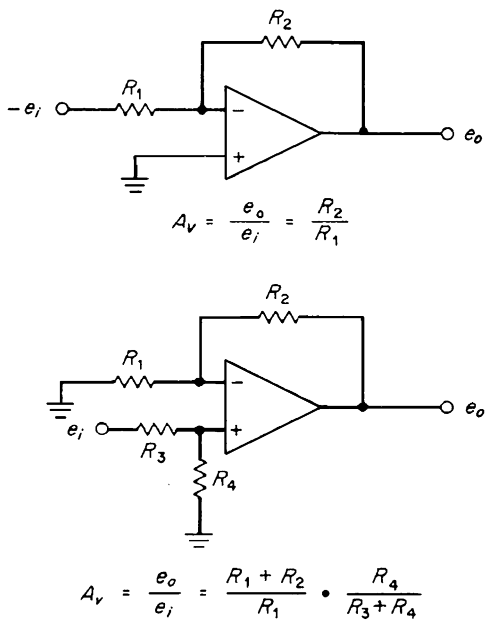


FIGURE 11.4.16 Basic operational-amplifier configurations.

There are many types of noise derived from a number of sources, but the types concerning amplifiers include $1/f$ or pink noise and thermal and other current- and voltage-associated effects within the amplifier. Pink noise is of concern at the very low end of the spectrum and involves dc amplifiers and some types of sensors. Generally the amplifier noise density can be expressed as follows:

$$E_n = G\sqrt{e_n^2 + (i_n R_{eq})^2 + e_t^2}$$

where e_n = op-amp noise voltage
i_n = op-amp current noise
R_{eq} = equivalent input source resistance (including feedback network)
e_t = thermal noise of source network (resistance)

and since

$$e_t = \sqrt{4KTR_{eq}}$$

where $K = 1.38 \times 10^{-23}$, $T = 300$ K, then

$$E_n = G\sqrt{e_n^2 + (i_n R_{eq})^2 + 4KTR_{eq}}$$

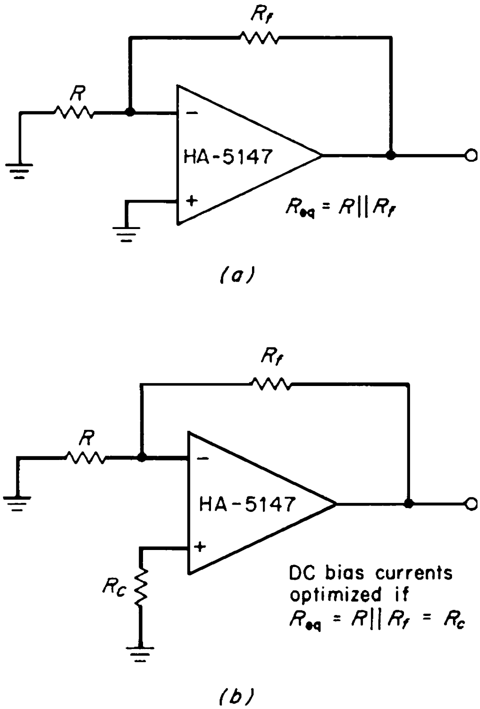


FIGURE 11.4.17 Simple op-amp configuration with and without bias.

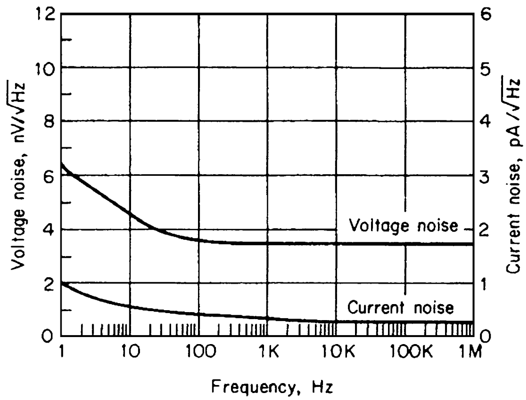


FIGURE 11.4.18 Voltage and current noise characteristics of a commercially available op amp.

The R_{eq} can become a predominant influence in the noise of the amplifier. This resistance is the combination of the resistive component of the input to the amplifier and the feedback network. It is obvious that the input to the amplifier should be kept as low as possible without loading down the device the amplifier is attached to. Figure 11.4.17 shows the R_{eq} of the simple amplifier. When offset bias is added to the op amp, the thermal and current noise components must be added to the noise equation. Usually the bias resistance is equal to the R_{eq} making the noise density

$$E_n = G\sqrt{e_n^2 + 2(i_n R_{eq})^2 + 8KTR_{eq}}$$

The amplifier design should have bandwidth restrictions on it so that the band extent does not go beyond that which is necessary. When the amplifier design is the first stage of a larger amplifier and it becomes the noise determining circuit, the bandpass restrictions do not have to be repeated further on.

There is one other consideration regarding the R_{eq} and the source resistances. If truly low noise is desired, noncarbon resistors are needed in this part of the circuit. Metal film resistors are best. Wire-would resistors are also a possibility, but not if high frequencies are involved. Carbon resistors have a larger "particle component" of noise beside the normal thermal noise.

Generally, the current noise i_n and voltage noise e_n remain relatively constant above 1 kHz and rise somewhat below 100 Hz, as shown in Fig. 11.4.18. If high values of R_{eq} or source resistances are necessary, op-amp selection is possible to reduce noise in parts of the spectrum. This is apparently a result of the biasing arrangement of different op amps with their input transistors ($e_n = f/\sqrt{i_c}$, $i_n = f\sqrt{i_c}$). With R_{eq} less than 1 kΩ, the thermal noise is expected to be low and the total noise is dominated by the voltage noise e_n.

POWER OPERATIONAL AMPLIFIERS

Clayton R. Roberts

Power Op Amps

The advent of the *power op amp* has provided us with another versatile design approach when high-current high-voltage drive requirements are needed. It is useful in motor drive circuits, deflection sweep circuits, many control functions, and high-quality audio amplifiers. Distortion levels below 0.05 percent can be achieved. In addition to the high-voltage high-current ratings the power op amps usually provide better matched components from internal control of the devices and laser trimming of the stages. Standard design approaches for op amps are used except for some considerations mentioned below. The definition of where the power op amp takes over from the standard op amp appears to be when the total supply voltage must be above 44 V and the delivered current is above 0.1 A.

Power Dissipation. In addition to the maximum voltage rating concerns, the power op-amp design must include operating temperature and dissipation values. Mounting of the device as well as the operating region of the parameters is a major design criterion. Figure 11.4.19 shows the effect of junction temperature of the power devices on failure rates as was derived from the base failure rate tables of MIL-HDBK-217C. Some power op-amp manufacturers design the cases so that they can be directly mounted to grounded heat sinks so that there is minimum thermal resistance to the sink. The required thermal resistance ϕ from a heat sink is calculated by the following:

$$\phi = (T_j - T_a)/P_{\mathrm{dis}} - \phi_{\mathrm{JC}}$$

where T_j = maximum junction temperature
T_a = maximum ambient temperature
P_{dis} = internal power dissipation
ϕ_{JC} = thermal resistance of the amplifier

Another heat-sink-design approach (Fig. 11.4.20) offers the approximate area required of a convection cooled surface.

The complexity of the power-dissipation calculation is very much dependent on the type of load that is being driven. Taking dc or instantaneous values of current and voltage, it may be expressed as

$$P_{\mathrm{dis}} = (V_s - V_o)I_o + V_{ss}(I_q)$$

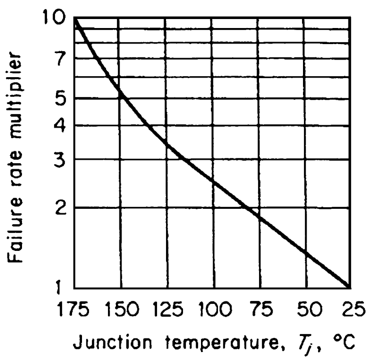


FIGURE 11.4.19 Failure rate (MTTF) vs. junction temperature.

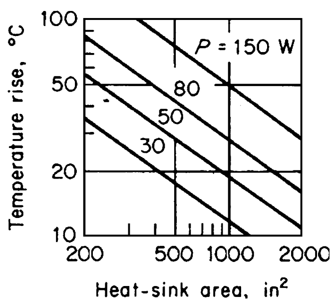


FIGURE 11.4.20 Heat-sink area estimated for temperature rise and several values of power dissipation.

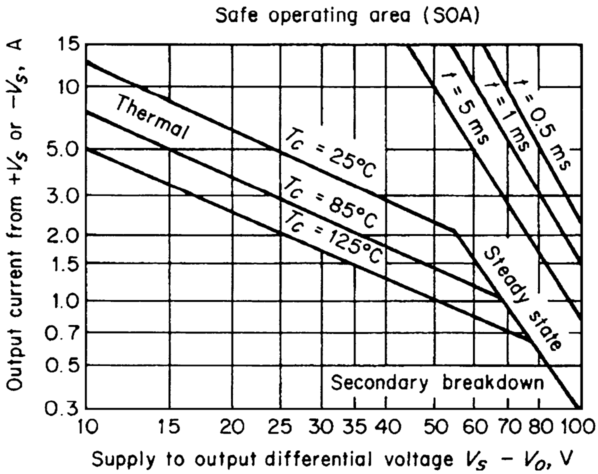


FIGURE 11.4.21 Safe operating curves for a power operational amplifier.

where V_s = one rail of the supply voltage
V_o = output voltage
I_o = output current
V_{ss} = rail-to-rail supply voltage
I_q = quiescent current

For sinusoidal waveforms the peak value of P_{dis} may be expressed as

$$P_{\text{dis}} = (V_s/2Z_l)[1 - \cos(60° - \theta)]$$

where θ is the absolute value of the phase angle of Z_l and Z is the magnitude of the load impedance. This function is approximate.

Safe-operating-area (SOA) curves are supplied by some manufacturers to aid in limiting the power dissipation to within specifications. Figure 11.4.21 shows a curve for a commercial power op amp. The secondary breakdown region shown results from excess current densities in the base region of the output devices when high collector current occurs simultaneously with high collector-emitter voltage. These curves include all contributors to the dissipation problem. The three curves at the right-hand end of Fig. 11.4.21 are the limits of the short-pulse power applications for the SOA curve. Pulse width values of 0.5, 1.0, and 5.0 ms are indicated. Additional foldover or limiting current protection can be obtained when operating near the limits (rail) (refer to manufacturers product data).

Output Loads. Output loads can have reactance components involved even if they are designed to be resistive. With the higher currents present, and especially transients, reactive, or flyback (reversed polarity) voltages can exceed specified limits. Most power op amps have built-in diodes between the two rails and the output to help suppress such transients. By necessity these can handle limited current. Several manufacturers recommend external diodes be placed between the output of the op amp and each of the supplies. The back emf must be considered in the design (SOA curves). This is especially true with loads such as motors. Start-up and reverse conditions must be accounted for. Some types of loudspeaker circuits present a similar situation.

Circuit Layout. Lead inductance must be kept to a minimum. This is sometimes taken care of by use of a microstrip-type low-impedance line approach. Power-supply transients are reduced to a minimum by bypassing right at the op amp as well as at the supply terminals. With high currents involved, bypass capacitors must have low reactance leads and several times their normal values. With low-distortion amplifiers, deteriorization

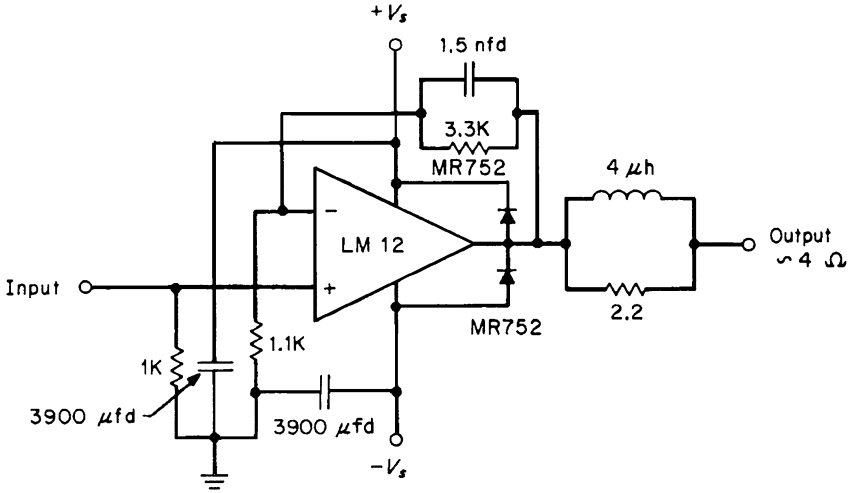


FIGURE 11.4.22 National Semiconductor LM12 power amplifier.

at the high-frequency end of the spectrum (higher distortion) can be the result of insufficient effective bypassing from the supply sources, but this is best taken care of by large short-lead bypasses at the device location. Ground loops and ground return interaction effects are also emphasized with the high-current devices. Input-output ground return coupling is to be avoided.

An audio power amplifier shown in Fig. 11.4.22 has some of the features indicated in a low-cost high-performance design. The LM12 (National Semiconductor) is a 150-W op amp capable of delivering that power to a 4-Ω load. The design includes Motorola MR752 diodes in its output to the rails. These diodes are capable of handling up to 400 A of surge current, 28 A average forward current, and a repetitive reverse voltage of 200 V. These specifications provide good protection against transients and back emf. The small capacitor in the feedback network aids in reducing the distortion at the high end of the spectrum and provides stability. The distortion is claimed to be about 0.01 percent. Good supply bypassing is obtained right at the device terminals. A central grounding point is used to avoid ground coupling of output to input. Thermal limiting, output current limiting, and over voltage shutdown is designed into the device. Note the parts count associated with circuit.